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# ENHANCED BER ANALYSIS AND MITIGATION OF ICI IN SFBC-OFDM SYSTEMS

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## ABSTRACT

SFBC-OFDM signals usage is advantageous in high-mobility broadband wireless access, where the channel is highly time as well as frequency-selective because of which the receiver experiences both ISI and ICI. ISI occurs due to the violation of the `quasi-static' fading assumption caused due to frequency- and/or time-selectivity of the channel. In addition, ICI occurs due to time-selectivity of the channel which results in loss of orthogonality among the subcarriers [1]. In this paper, we are concerned with the detection of SFBC-OFDM signals on time and frequency-selective MIMO channels. In this work, we investigate interference cancellation techniques to mitigate the effects of ICI induced by time-selectivity and ISI induced by frequency-selectivity of the channel. This paper is organized as follows. In Section I, we investigated the enhanced performance analysis of SFBC-OFDM system. In Section II, we presented the SFBC-OFDM system model. In Section III, we presented the proposed interference cancellation algorithm for the considered SFBC-OFDM system. Results and discussions are presented in Section IV. Conclusion is given in Section V.

**KEYWORDS:** MIMO Channels, SFBC-OFDM Signals, Broadband Wireless Access

## I. INTRODUCTION

## PERFORMANCE OF SPACE-FREQUENCY BLOCK CODES

In this section, SFBC transmit diversity technique is applied into the OFDM system. Simply, the 2Tx/1Rx and 2Tx/2Rx antenna configurations are considered to compare the system performance of the MIMO OFDM system. Here, we discuss the traditional MIMO SFBC-OFDM structure with 2Tx/1Rx and 2Tx/2Rx antennas. In this section, SFBC has been with different digital modulation schemes such as BPSK and QPSK modulation techniques. Space Frequency Coding (SFC) is an efficient approach to exploit the enormous diversity offered by the MIMO. It is used to obtain gains due to spatial diversity via multiple transmit and receive antennas. Moreover, a diversity gain proportional to the number of antennas at both transmit and receive sides can be achieved. One popular representation of these codes is the Alamouti's scheme [2] for two transmit antennas.

### 1.1 Alamouti's Scheme with 2Tx-1Rx Antennas

Alamouti introduced a very simple scheme of SFBC allowing transmissions from two antennas with the same data rate as on a single antenna, but increasing the diversity at the receiver from one to two in a Rayleigh fading channel. As shown in Figure 1, the Alamouti's algorithm uses the space and frequency domain to encode data with increase in the performance of the system by coding the signals over the different transmitter branches [1]. Thus, the Alamouti's code achieves diversity two with full data rate as it transmits two symbols in two frequency slots. In the first frequency slot, transmit antennas Tx1 and Tx2 are sending symbols $S_0$ and $S_1$, respectively. In the next frequency slot, symbols $-S_1^*$ and

$s_0^*$ are sent, where $(\cdot)$*denotes complex conjugation. Furthermore, it is supposed that the channel, which has transmission coefficients $h_1$ and $h_1$, remains constant and frequency is flat over the two consecutive time steps. The received vector, R, is formed by stacking two consecutively received data samples in frequency, resulting in

$$R = Sh + w \tag{1.1}$$

$where\ R = [r_0, r_1]^T$ Represents the received vector [186],

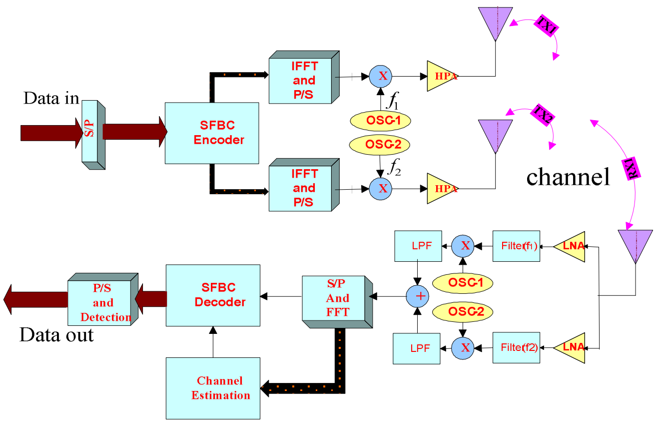


**Figure 1: 2 X 1 SFBC-OFDM Transceiver**

$h = [h_1, h_2]^T$ Is the complex channel vector, $w = [w_1, w_2]^T$ is the noise at the receiver and S defines the SFBC [27],

$$S = \begin{pmatrix} s_0 & s_1 \\ -s_1^* & s_0^* \end{pmatrix} \tag{1.2}$$

The vector equation in (4.30) can be read explicitly as

$$\begin{aligned} r_0 &= s_0 h_1 + s_1 h_2 + w_0 \\ r_1 &= -s_1^* h_1 + s_0^* h_2 + w_1 \end{aligned} \tag{1.3}$$

At the receiver, the vector **y** of the received signal is formed according to $y = \left[r_0, r_1^*\right]^T$. The above eqn. (4.32) can be rewritten in a matrix system as

$$\begin{pmatrix} r_0 \\ r_1^* \end{pmatrix} = \begin{pmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{pmatrix} \begin{pmatrix} s_0 \\ s_1 \end{pmatrix} + \begin{pmatrix} w_0 \\ w_1^* \end{pmatrix} \tag{1.4}$$

The short notation for this system is the following:

$$y = H_v s + \tilde{w} \tag{1.5}$$

Where $\tilde{w}$ represents the new noise vector obtained after the conjugation of the second equation, $\tilde{w} = \left[w_0, w_1^*\right]^T$ the estimated transmitted signal is then calculated from the formula

$$\tilde{s} = H_V^H y \quad where \quad y = \left[r_0, r_1^*\right]^T$$

$$\begin{aligned} r_0 &= s_0 h_1 + s_1 h_2 + w_0 \\ r_1^* &= s_0 h_2^* - s_1 h_1^* + w_1^* \end{aligned} \tag{1.6}$$

$where \quad H_V^H = \begin{pmatrix} h_1^* & h_2 \\ h_2^* & -h_1 \end{pmatrix}$ Is Hermitian of the virtual channel matrix? $H_V$

$$\tilde{s} = \begin{pmatrix} \tilde{s}_0 \\ \tilde{s}_1 \end{pmatrix} = H_V^H H_V \begin{pmatrix} s_0 \\ s_1 \end{pmatrix} + H_V^H \begin{pmatrix} w_0 \\ w_1^* \end{pmatrix} \tag{1.7}$$

The resulting virtual (2 × 2) channel matrix $H_V$ is orthogonal, i.e.

$$H_V^H H_V = H_V H_V^H = h^2 I_2 \tag{1.8}$$

Due to this orthogonality, the Alamouti's scheme decouples the MISO channel into two virtually independent channels with channel gain $h^2$ and diversity d = 2. The channel gain specifies that transmitted symbols can be estimated at the receiver as the result of multiplying the received signals by the Hermitian of the virtual channel matrix. After performing the corresponding operations it results in a signal with a gain of $h^2$ plus some modified noise.

$$\hat{s} = h^2 I_2 s + \hat{w} \tag{1.9}$$

Where, $s$ is the transmitted signal, $I_2$ is the 2×2 identity matrix, $h^2 = |h_1|^2 + |h_2|^2$ and $\hat{w} = \begin{pmatrix} h_1^* w_0 + h_2 w_1^* \\ h_2^* w_0 - h_1 w_1^* \end{pmatrix}$ is modified noise.

### 1.2 Simulation Results

The plot shown in Figure 2 shows the BER comparative performance of Alamouti's SFBC-OFDM in different channel situations. The 2 TX and 1 RX SFBC-OFDM with Rayleigh channel shows outperformance than standard OFDM system.

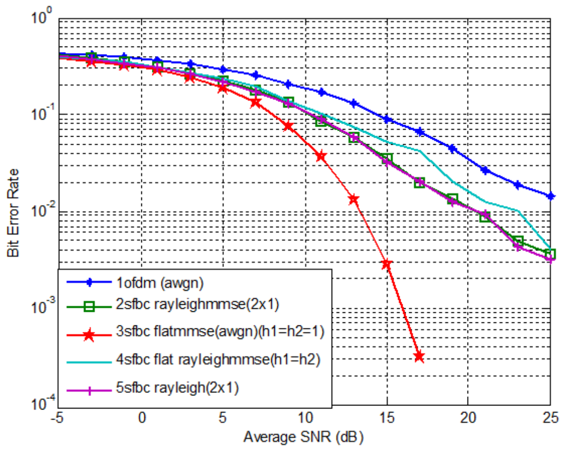


**Figure 2: BER Performance of 2 X 1 Alamouti SFBC-OFDM**

### 1.3 Alamouti's with 2Tx - 2Rx Antennas

Systems with two transmit antennas and two receive antennas as the one shown in Figure 3, is analyzed next. The already explained steps are applied to each of the receive antennas, denoting the received signal in the first and second time slot as $\mathbf{r}_1$ and $\mathbf{r}_2$, respectively [4].

The received signal from a 2 × 2 Alamouti's scheme, as depicted above, is

$$y=\begin{pmatrix} r_0(1) \\ r_0(2) \\ r_1^*(1) \\ r_1^*(2) \end{pmatrix}=\begin{pmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \\ h_{21}^* & -h_{11}^* \\ h_{22}^* & -h_{12}^* \end{pmatrix}\begin{pmatrix} s_0 \\ s_1 \end{pmatrix}+\begin{pmatrix} w_0(1) \\ w_0(2) \\ w_1^*(1) \\ w_1^*(2) \end{pmatrix} \tag{1.10}$$

The estimated transmitted signal can be calculated from $\hat{s}=H_V^H y$ *where* $y=\left[r_0(1), r_0(2), r_1^*(1), r_1^*(2)\right]^T$ and $(.)^H$ represents the Hermitian operation

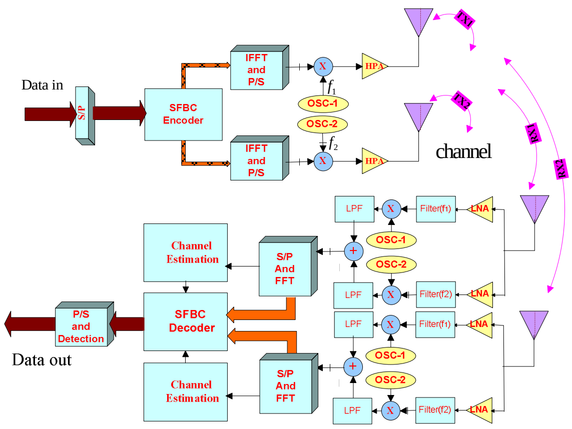


**Figure 3: 2 x 2 SFBC-OFDM Transceiver**

The virtual channel matrix, is expressed as $H_V$ is expressed as

$$H_V=\begin{pmatrix} h_{11} & h_{21} \\ h_{12} & h_{22} \\ h_{21}^* & -h_{11}^* \\ h_{22}^* & -h_{12}^* \end{pmatrix} \quad \textit{Therefore} \quad H_V^H=\begin{pmatrix} h_{11}^* & h_{12}^* & h_{21} & h_{22} \\ h_{21}^* & h_{22}^* & -h_{11} & -h_{12} \end{pmatrix} \tag{1.11}$$

The obtained result for the process of estimating the transmitted symbols is

$$\hat{s}=h^2 I_2 s+\hat{w} \tag{1.12}$$

Where, $I_2$ is the 2×2 identity matrix, **s** is the transmitted Signal, $h^2=|h_{11}|^2+|h_{21}|^2+|h_{12}|^2+|h_{22}|^2$ is the power gain of the channel[3], and

$$\hat{w}=\begin{pmatrix} h_{11}^* w_0(1)+h_{12}^* w_0(2)+h_{21} w_1^*(1)+h_{22} w_1^*(2) \\ h_{21}^* w_0(1)+h_{22}^* w_0(2)-h_{11} w_1^*(1)-h_{12} w_1^*(2) \end{pmatrix}$$ Represents some modified noise.

## 1.4 Simulation Results

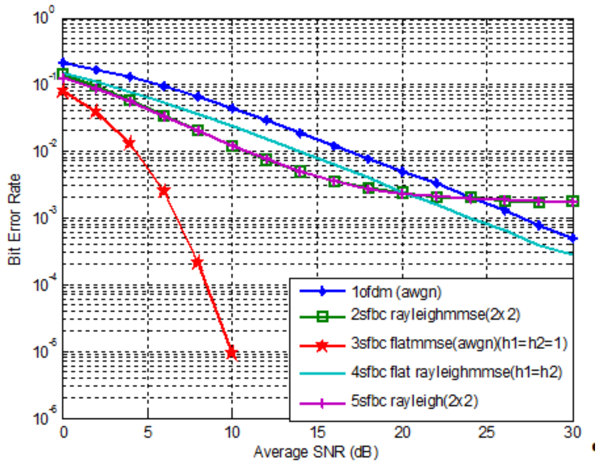


**Figure 4: BER Performance of 2 X 2 Alamouti SFBC-OFDM**

The plot showed in Figure 4.shows the BERs comparative performance of Alamouti's SFBC-OFDM in different channel situations. The 2 TX and 2 RX SFBC-OFDM with Rayleigh channel shows outperformance than standard OFDM system for low SNR values. From the Figure 5, we can see the performance of space-frequency block codes using QPSK, and BPSK modulation schemes. The BER performance of the system using BPSK modulation is better than the BER performance of space-time block codes using QPSK modulation by approximately 3 to 4 dB.

The performance of space-time block codes using two transmitters and two receivers shows better performance than that of space-time block codes using two transmitters and one receiver antenna by approximately 7 to 8 dB. From the plots, it can be seen that the BER reduces with the number of transmit and receive antennas and BER increases with the increase in modulation order.

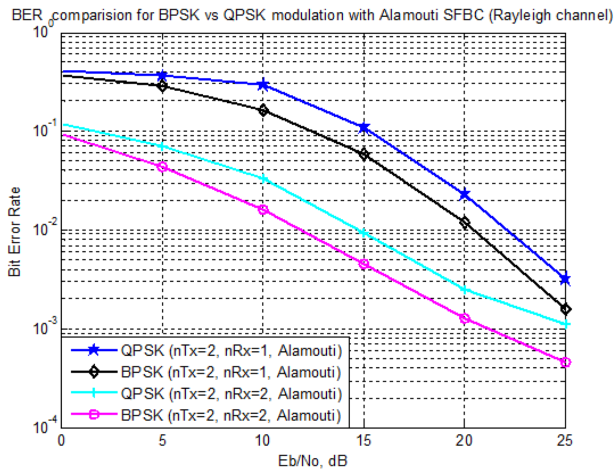


**Figure 5: BER Performance of Alamouti's SFBC-OFDM (BPSK Vs QPSK)**

## 1.5 CONCLUSIONS

The combination of OFDM with SFBC allowed the creation of codes known as SFBC-OFDM offering low decoding complexity and bandwidth efficiency as realized in SFBC for single carrier systems.

STBC-OFDM and SFBC-OFDM systems have been adopted by standard such as WiMax and 4G and have been under extensive investigation by researchers.

In this section, a comprehensive investigation of OFDM and in particular MIMO-OFDM systems was conducted. Two coding schemes, STBC-OFDM and SFBC-OFDM, have been described in detail and simulation results have been presented for different number of transmit and receive antennas and under various modulation schemes. In the next section, we are concerned with the detection of SFBC-OFDM signals on time and frequency-selective MIMO channels. Specifically in section III, we proposed and evaluated the performance of an interference cancelling receiver for SFBC-OFDM which effectively alleviates the effects of ISI and ICI.

## II. ICI AND ISI MITIGATION IN SFBC-OFDM SYSTEMS

### 2.1 SFBC-OFDM System Model

MIMO-OFDM system with $N_c$ sub carriers, $N_t$ transmit antennas and $N_r$ receiving antennas [5].

- $X_k(i)$ Denotes the complex data symbol transmitted on the $k^{th}$ subcarrier of an OFDM symbol from the ith transmit antenna. The symbols are transmitted in parallel on $N_c$ sub carriers using $N_t$ transmit antennas.
- To generate these parallel sub-carriers in OFDM, an IDFT/IFFT is applied to a block of '*L*' data symbols.
- To avoid ISI due to Channel delay spread, CP symbols are inserted into the block.
- Introduction of CP samples eliminates ISI and converts the linear convolution between the transmit symbols and the channel to a circular convolution.
- Next, these CP samples are removed at the receiver.
- Finally, A DFT/FFT is used at the receiver to recover the block of '*L*' received symbols.

#### 2.1.1 Derivation of ICI Coefficients

The Discrete-time sequence transmitted by the $i^{th}$ transmit antenna, after IDFT processing and insertion of guard interval of $n_g$ samples is

$$x_n^{(i)} = \frac{1}{N_c}\sum_{k=0}^{N_c-1} X_{k+1}^{(i)} e^{j\frac{2\pi nk}{N_c}} \qquad -n_g \le n \le N_c - 1 \tag{2.1}$$

Assuming perfect carrier synchronization, timing, and sampling at the RX , the discrete-time received sequence at the $j^{th}$ receive antenna, $j=1,2\ldots\ldots\ldots\ldots N_r$ is

$$y_n^{(j)} = \sum_{i=1}^{N_t}\sum_{l=0}^{L-1} h^{(i,j)}(n;l)x_{(n-l)}^{(i)} + w_n^{(j)},\ j = 1,\ldots\ldots.N_r;$$

$$-n_g \le n \le N_c - 1$$

Here $h^{(i,j)}\ (n;l)$ is the discrete-time, time-varying (time-selective) 'L'-length (frequency selective) channel

impulse response between the $i^{th}$ transmit and $j^{th}$ receive antennas.

$w_n^{(j)} \rightarrow$ Additive noise on the $j^{th}$ receive antenna.

$$H_k^{(i)} = \sum_{n=0}^{L-1} h_n^{(i)} e^{j\frac{2\pi nk}{N}} \qquad H(p) = \sum_{i=1}^{N_t} h^{(i,j)}(n;p)$$

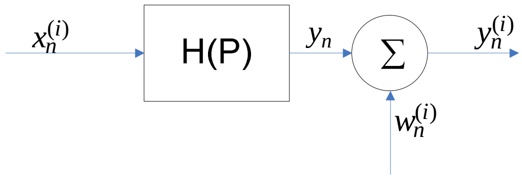


**Figure 6: Receiver Signal Model**

$$y_n^{(j)} = H(p) * x_n^{(i)} + w_n^{(j)}$$

$$y_n^{(j)} = \sum_{l=0}^{L-1} H(L) x_{(n-l)}^{(i)} + w_n^{(j)}$$

$$y_n^{(j)} = \sum_{l=0}^{L-1} \sum_{i=1}^{N_t} h^{(i,j)}(n;l)\ x_{(n-l)}^{(i)} + w_n^{(j)}$$

$$j = 1, \ldots\ldots\ldots N_r \quad ,$$
$$-n_g \leq n \leq N_c - 1$$

$$\boxed{y_n^{(j)} = \sum_{i=1}^{N_t} \sum_{l=0}^{L-1} h^{(i,j)}(n;l)\ x_{(n-l)}^{(i)} + w_n^{(j)}} \tag{2.2}$$

After guard interval removal and DFT operation, the received signal on the $k^{th}$ subcarrier on the $j^{th}$ receive antenna

$$Y_k^{(j)} \text{ is } Y_k^{(j)} = \sum_{n=0}^{N_c-1} y_n^{(j)}\ e^{-j\frac{2\pi nk}{N_c}}$$

$$= \sum_{n=0}^{N_c-1} \left\{ \sum_{i=1}^{N_t} \sum_{l=0}^{L-1} h^{(i,j)}(n;l)\ x_{(n-l)}^{(i)} + w_n^{(j)} \right\} e^{-j\frac{2\pi nk}{N_c}}$$

$$= \sum_{n=0}^{N_c-1} \sum_{i=1}^{N_t} \sum_{l=0}^{L-1} h^{(i,j)}(n;l)\ x_{(n-l)}^{(i)} e^{-j\frac{2\pi nk}{N_c}} + \underbrace{\sum_{n=0}^{N_c-1} w_n^{(j)} e^{-j\frac{2\pi nk}{N_c}}}_{W_k^{(j)}}$$

$$we\ know\ that\quad x_n^{(i)} = \frac{1}{N_c}\sum_{k=0}^{N_c-1} X_{k+1}^{(i)} e^{j\frac{2\pi nk}{N_c}} = \frac{1}{N_c}\sum_{m=0}^{N_c-1} X_{m+1}^{(i)} e^{j\frac{2\pi nm}{N_c}}$$

$$x_{(n-l)}^{(i)} = \frac{1}{N_c}\sum_{m=0}^{N_c-1} X_{m+1}^{(i)} e^{j\frac{2\pi(n-l)m}{N_c} i\theta}$$

$$x_{(n-l)}^{(i)} = \frac{1}{N_c}\sum_{m=0}^{N_c-1} X_{m+1}^{(i)}\ e^{j\frac{2\pi nm}{N_c}}\ e^{-j\frac{2\pi lm}{N_c}}$$

Hence

$$Y_k^{(j)} = \sum_{i=1}^{N_t}\frac{1}{N_c}\sum_{n=0}^{N_c-1}\sum_{l=0}^{L-1} h^{(i,j)}(n;l)\left\{\sum_{m=0}^{N_c-1} X_m^{(i)}\ e^{j\frac{2\pi nm}{N_c}}\ e^{-j\frac{2\pi lm}{N_c}}\right\} e^{-j\frac{2\pi nk}{N_c}} + W_k^{(j)}$$

$$= \frac{1}{N_c}\sum_{i=1}^{N_t}\sum_{n=0}^{N_c-1}\sum_{l=0}^{L-1} h^{(i,j)}(n;l)\Big[\sum_{\substack{only\\ m=k}} X_{m=k}^{(i)} e^{j\cancel{\frac{2\pi nk}{N_c}}}\ e^{-j\frac{2\pi lk}{N_c}}\ e^{-j\cancel{\frac{2\pi nk}{N_c}}}$$

$$+\sum_{\substack{m\neq k\\ m=0}}^{N_c-1} X_m^{(i)}\ e^{j\frac{2\pi nm}{N_c}}\ e^{-j\frac{2\pi lm}{N_c}}\ e^{-j\frac{2\pi nk}{N_c}}\Big] + W_k^{(j)}$$

*Assuming*

$$\frac{1}{N_c}\sum_{n=0}^{N_c-1}\sum_{l=0}^{L-1} h^{(i,j)}(n;l)\quad e^{j\frac{2\pi n(m-k)}{N_c}}\quad e^{-j\frac{2\pi lm}{N_c}} = G_{k,m}^{(i,j)}\quad for\ \ k\neq m.$$

*and*

$$\frac{1}{N_c}\sum_{n=0}^{N_c-1}\sum_{l=0}^{L-1} h^{(i,j)}(n;l)\qquad e^{-j\frac{2\pi lm}{N_c}} = G_{k,k}^{(i,j)}\quad i.e,\ k=m.$$

Then

$$Y_k^{(j)} = \sum_{i=1}^{N_t} G_{k,k}^{(i,j)} X_k^{(i)} + \underbrace{\sum_{i=1}^{N_t}\sum_{\substack{m\neq k\\ m=0}}^{N_c-1} G_{k,m}^{(i,j)} X_m^{(i)}}_{ICI} + \underbrace{W_k^{(j)}}_{\downarrow\ Noise} \tag{2.3}$$

Where the coefficients $G_{k,m}^{(i,j)}$ are given by

$$G_{k,m}^{(i,j)} = \frac{1}{N_c}\sum_{n=0}^{N_c-1}\sum_{l=0}^{L-1} h^{(i,j)}(n;l) e^{\frac{j2\pi n(m-k)}{N_c}}\ e^{\frac{-j2\pi ml}{N_c}} \tag{2.4}$$

$G_{k,m}^{(i,j)}$ , $k\neq m$ Denotes the amount of carrier leak from $m^{th}$ subcarrier to the $k^{th}$ subcarrier, which contributes to the ICI term in eqn. (2.3). If the channel is not time-selective i.e., time-flat between all transmit-receive antenna pairs. i.e., if

$$h^{(i,j)}(n_1;l) = h^{(i,j)}(n_2;l)\quad \forall\ i,j,l\quad -n_g \le n_1, n_2 \le N_c - 1 \tag{2.5}$$

then $G_{k,m}^{(i,j)} = 0$ for $k\neq m$ and eqn. (2.3) reduces to (ICI term=0)

$$Y_k^{(j)} = \sum_{i=1}^{N_t} \tilde{G}_{k,k}^{(i,j)} \; X_k^{(i)} + W_k^{(j)} \tag{2.6}$$

Where $\tilde{G}_{k,k}^{(i,j)}$ is $G_{(k,k)}^{(i,j)}$ for time-flat channels, given by

$$\tilde{G}_{k,k}^{(i,j)} = \frac{1}{N_c} \sum_{l=0}^{L-1} h^{(i,j)}(n_1, l) \; e^{-j\frac{2\pi kl}{N_c}} \tag{2.7}$$

## III. PROPOSED IC RECEIVER FOR SFBC-OFDM WITH CARRIER FREQUENCY OFFSET

### 3.1 Space-Frequency Block Coded OFDM

In this section, we adopt the above system model to the case of space-frequency coded symbols transmitted on different transmit antennas (space) and different subcarriers (frequency). Let, $X_k^{(i)}$ are the symbols obtained from the space-frequency coding scheme. Specifically, we consider the use of OSTBCs as the space-frequency codes. Let $K$ denote the length of one space-frequency code block. We divide $N_c$ subcarriers into $N_g$ groups each having $K$ subcarriers so that $N_c = N_g K + \kappa$ where each group is called one SFC block. For example, $K = 2$ for Alamouti code. If $N_c$ is not a multiple of $K$, then there will not be any transmission on $\kappa$ subcarriers, or, alternatively, these $\kappa$ subcarriers can be used for pilot transmission.

The SFBC-OFDM frame thus obtained can be written as an $N_c \times N_t$ matrix,

$$X = \begin{bmatrix} X_1^{(1)} & \cdots & X_1^{(N_t)} \\ X_2^{(1)} & \cdots & X_2^{(N_t)} \\ \vdots & \vdots & \vdots \\ X_{N_c}^{(1)} & \cdots & X_{N_c}^{(N_t)} \end{bmatrix} = \begin{bmatrix} X(1) \\ \vdots \\ X(N_g) \\ 0^{k \times N_t} \end{bmatrix} \tag{3.1}$$

Where $X(q)$ is a $K \times N_t$ matrix, $q = 1, 2, \ldots, N_g$. $X(q)$ is from a OSTBC in $P$ complex information symbols $[v_1(q), v_2(q), \ldots, v_P(q)]$ and rate-$P/K$. The $i^{th}$ column of X is transmitted on the $i^{th}$ transmit antenna after IDFT processing. For example, for SFBC

OFDM with $N_c = 4$ using Alamouti code, $N_t = 2$ and the matrix X can be written as

$$X_{Alamouti} = \begin{bmatrix} v_1(1) & v_2(1) \\ -v_2(1)^* & v_1(1)^* \\ v_1(2) & v_2(2) \\ -v_2(2)^* & v_1(2)^* \end{bmatrix} \tag{3.2}$$

By stacking all the $K$ rows of $X(q)$ into one $KN_t \times 1$ vector, $\overline{x}(q)$. This $\overline{x}(q)$ vector can be written as

$$\overline{x}(q) = Av(q) \tag{3.3}$$

Where $v(q)$ is a $2P \times 1$ vector given by

$$v(q) = [v_{1I}(q), \cdots, v_{PI}(q), v_{1Q}(q), \cdots, v_{PQ}(q)] \tag{3.4}$$

Where $v_{pI}(q)$ and $v_{pQ}(q)$, respectively, are the real and imaginary parts of the $p^{th}$ complex information symbol in the $q^{th}$ group, $p = 1, 2, \ldots, P$, $q = 1, 2, \ldots, N_g$. The matrix A in (3.3) is a $KN_t \times 2P$ complex matrix which performs the coding on $v(q)$. For the SFBC-OFDM with Alamouti's code, A is given by

$$A_{Alamouti} = \begin{bmatrix} 1 & 0 & j & 0 \\ 0 & 1 & 0 & j \\ 0 & -1 & 0 & -j \\ 1 & 0 & -j & 0 \end{bmatrix} \tag{3.5}$$

Using eqn. (3.5) in eqn. (3.3), we get

$$\bar{x}_{Alamouti}(q) = [v_1(q), v_2(q), -v_2(q)^*, v_1(q)^*]^T \tag{3.6}$$

And

$$X_{Alamouti}(q) = \begin{bmatrix} v_1(q) & v_2(q) \\ -v_2(q)^* & v_1(q)^* \end{bmatrix} \tag{3.7}$$

**3.2 Received Signal for SFBC-OFDM**

At the receiver, the DFT outputs, $Y_k^{(j)}$'s, of eqn. (6.3) are stacked to form a $KN_r \times 1$ vector for each group, as

$$y(q) = [Y_{r+1}^{(1)}, \ldots, Y_{r+1}^{(N_r)}, \ldots, Y_{r+k}^{(1)}, \ldots, Y_{r+k}^{(N_r)}]^T \tag{3.8}$$

Where $r = (q - 1) K$. Now, $y(q)$ in eqn. (6.25) can be written in the form

$$y(q) = \bar{H}(q)\bar{x}(q) + s(q) + w(q) \tag{3.9}$$

Where $\bar{H}(q)$ is a $KN_r \; X \; KN_t$ block diagonal matrix, given by

$$\bar{H}(q) = \begin{bmatrix} H(q,1) & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & H(q,k) \end{bmatrix} \tag{3.10}$$

Where $H(q, s)$ is the $N_r \times N_t$ channel matrix, $q = 1, 2, \ldots, N_g$, $s = 1, 2, \ldots, K$,

Given by

$$H(q,s) = \begin{bmatrix} G_{u,u}^{(1,1)} & \cdots & G_{u,u}^{(1,N_t)} \\ \vdots & \ddots & \vdots \\ G_{u,u}^{(N_r,1)} & \cdots & G_{u,u}^{(N_r,N_t)} \end{bmatrix} \tag{3.11}$$

Where $u = (q - 1) K + s$. The noise vector w $(q)$ in eqn. (3.9) is given by

$$w(q) = [W_{r+1}^{(1)}, \cdots, W_{r+1}^{(N_r)}, \cdots, W_{r+K}^{(1)}, \cdots, W_{r+K}^{(N_r)}]^T \tag{3.12}$$

Where $r = (q - 1) K$. The interference vector $s(q)$ in eqn. (3.9) can be written in the form

$$s(q) = \bar{R}(q)\bar{x}(q) + \sum_{b=1, b\neq q}^{N_g} \bar{Q}(q,b)\bar{x}(b) \tag{3.13}$$

Where the block matrices $\bar{R}(q)$ and $\bar{Q}(q,b)$ are given by

$$\bar{R}(q) = \begin{bmatrix} 0 & R_{q'+1,q'+2} & \cdots & R_{q'+1,q'+K} \\ R_{q'+2,q'+1} & 0 & \cdots & R_{q'+2,q'+K} \\ \vdots & \vdots & \ddots & \vdots \\ R_{q'+K,q'+1} & R_{q'+K,q'+2} & \cdots & 0 \end{bmatrix} \tag{3.14}$$

$$\bar{Q}(q,b) = \begin{bmatrix} R_{q'+1,b'+1} & \cdots & R_{q'+1,b'+K} \\ \vdots & \ddots & \vdots \\ R_{q'+K,b'+1} & \cdots & R_{q'+k,b'+K} \end{bmatrix} \tag{3.15}$$

Where, $q' = (q-1)K, \quad b' = (b-1)K, \ and$

$$R_{x,y} = \begin{bmatrix} G_{x+1,y+1}^{(1,1)} & \cdots & G_{x+1,y+K}^{(1,N_t)} \\ \vdots & \ddots & \vdots \\ G_{x+K,y+1}^{(N_r,1)} & \cdots & G_{x+K,y+K}^{(N_r,N_t)} \end{bmatrix} \tag{3.16}$$

### 3.3 Detection of SFBC-OFDM

For the case of time-flat and frequency-flat conditions (i.e., no ISI and no ICI), the detector for OSTBCs presented in [6] for conventional space-time codes can be used for detecting SFBC-OFDM as follows. For the time-flat case, eqn. (3.9) can be written as

$$y(q) = H_{eq}(q)v(q) + w(q) \tag{3.17}$$

Where $H_{eq}(q)$ is a $KN_r \times 2P$ equivalent channel matrix for the $q^{th}$ group, given by

$$H_{eq}(q) = \bar{H}(q)A \tag{3.18}$$

For the frequency-flat case, the quasi-static assumption holds, i.e., in eqn. (3.10)

$$H(q,1) = H(q,2) = \cdots = H(q,K), \quad \forall q \tag{3.19}$$

Now, from [190], the optimal detector for SFBC-OFDM under the above conditions can be shown to be of the form

$$\hat{y}(q) = \Re(H_{eq}^{*}(q)y(q)) \tag{3.20}$$

Where $\hat{y}(q)$ is a $2K \times 1$ vector containing the estimates of the real and imaginary parts of the complex information symbols in a stacked up fashion, which can be shown to be

$$\begin{aligned} \hat{y}(q) &= \Re[H_{eq}^{*}(q)H_{eq}(q)]V(q) + \Re[H_{eq}^{*}(q)w(q)] \\ &= \Lambda(q)V(q) + \hat{w}(q) \end{aligned} \tag{3.21}$$

Where $\wedge(q) = \Re\left[H_{eq}^{*}(q)H_{eq}(q)\right]$ is a diagonal matrix, and hence there will not be any inter-symbol interference. Furthermore, it can also be shown that under these conditions the noise term, $\hat{w}(q) = \Re\left[H_{eq}^{*}(q)w(q)\right]$ is white Gaussian. Hence, the Euclidean distance based symbol-by-symbol detection on $\hat{y}(q)$ is optimal. The optimum detector for this system would be a ML detector in *P* variables which results in exponential receiver complexity. A time-selective channel in this case results in inter-carrier interference eqn. (3.9). We further observe that cancellation techniques can be employed to recover the performance loss due to time- and frequency-selectivity induced interferences, which is our focus in the following section.

### 3.4 ISI and ICI Cancellation Using Interference Cancellation (IC) Algorithm

In this section, we propose a novel two-step PIC detector that cancels ISI and ICI in SFBC-OFDM. The proposed detector estimates and cancels the ISI (caused due to the violation of the quasi-static assumption) in the first step, and then estimates and cancels the ICI (caused due to loss of subcarrier orthogonality) in the second step. This two step procedure is then carried out in multiple stages. The proposed detector is presented in the following:

We consider perfect channel knowledge at the receiver. So, in the notation, we will not differentiate between the actual channel and the channel estimate available at the receiver. The detector, however, can work with imperfect channel estimates. First, we model the ISI caused by the violation of the quasi-static assumption. To do that, we split the block diagonal matrix $\bar{H}(q)$ in eqn. (6.34) into two parts: i) a quasi-static part $\bar{H}_{qs}(q)$ and ii) a non-quasi-static part $\bar{H}_{nqs}(q)$, such that

$$\bar{H}(q) = \bar{H}_{qs}(q) + \bar{H}_{nqs}(q) \tag{3.22}$$

$$\bar{H}_{qs}(q) = \begin{bmatrix} H(q,1) & 0 & \cdots & 0 \\ 0 & H(q,1) & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & H(q,1) \end{bmatrix} \tag{3.23}$$

$$\bar{H}_{nqs}(q) = \begin{bmatrix} 0 & 0 & \cdots & 0 \\ 0 & \Delta H(q,2) & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & \Delta H(q,K) \end{bmatrix} \tag{3.24}$$

*Where*

$$\Delta H(q,m) = H(q,m) - H(q,1)$$

Similarly, we split the equivalent channel matrix $H_{eq}(q)$ ,

$$H_{eq}(q) = H_{eq-qs}(q) + H_{eq-nqs}(q) \tag{3.25}$$

Where

$$H_{eq-qs}(q) = \bar{H}_{qs}(q)A, \text{ and } H_{eq-nqs}(q) = \bar{H}_{nqs}(q)A$$

Based on the above formulations and eqn. (3.4), we can write eqn. (3.17) as

$$y(q) = \bar{H}_{qs}(q)\bar{x}(q) + \underbrace{\bar{H}_{nqs}(q)\bar{x}(q)}_{QS\ violation} + \underbrace{\bar{R}(q)\bar{x}(q) + \sum_{b=1,b\neq q}^{N_g} \bar{Q}(q,b)\bar{x}(b) + w(q)}_{loss\ of\ \ orthogonality} \tag{3.26}$$

As in eqn. (3.20), an estimate of *y (q)* can be obtained as

$$\begin{aligned}\hat{y}(q) &= \Re(H^*_{eq-qs}(q)y(q)) \\ &= \underbrace{\Re[H^*_{eq-qs}(q)H_{eq-qs}(q)]v(q)}_{desired\ \ signal} + \underbrace{\Re[H^*_{eq-qs}(q)H_{eq-nqs}(q)]v(q)}_{ISI} \\ &\quad \underbrace{+\Re[H^*_{eq-qs}(q)\bar{R}(q)\bar{X}(q)] + \Re[H^*_{eq-qs}(q)\sum_{b=1,b\neq q}^{N_g}\bar{Q}(q,b)\bar{x}(q)}_{ICI} \\ &\quad \underbrace{+\Re[H^*_{eq-qs}(q)w(q)}_{noise}\end{aligned} \tag{3.27}$$

As can be seen, eqn. (3.27) the estimate $\hat{y}(q)$ contains the desired signal, ISI, ICI, and noise components. Based on this received signal model in eqn. (3.27) and the knowledge of the matrices $H_{eq-qs}(q)$, $H_{eq-nqs}(q)$, $\bar{R}(q)$ and $\bar{Q}(q,b)\forall q,b$, we formulate the proposed interference estimation and cancellation procedure as follows.

- For each SF code block q, estimate the information symbols $\hat{V}(q)$ from eqn. (3.27), ignoring ISI and ICI.
- For each SF code block *q*, obtain an estimate of the ISI (i.e., an estimate of the 2nd term in eqn. (3.27)) from the estimated symbols $\hat{v}(q)$ in the previous step.
- Cancel the estimated ISI from $\hat{y}(q)$.
- Using $\hat{v}(q)$ from step 1, regenerate $\hat{\bar{x}}(q)$ using eqn. (3.3). Then, using $\hat{\bar{x}}(q)$, obtain an estimate of the ICI (i.e., an estimate of the sum of 3rd and 4th terms in eqn. (3.27)).
- Cancel the estimated ICI from the ISI cancelled output in step 3.
- Take the ISI and ICI cancelled output from step 5 as the input, and go back to step 1 (for the next stage of cancellation).

Based on the above procedure steps and defining $\wedge(q) = \Re\left[H^*_{eq-qs}(q)H_{eq-qs}(q)\right]$, the cancellation algorithm for the $m^{th}$ stage can be summarized as follows.

Initialization: Set m=1.

Evaluate

$$\hat{y}^{(m)}(q) = \Re(H^*_{eq-qs}y(q)), \quad 1 \leq q \leq N_g \tag{3.28}$$

Loop $z_1$

Estimate

$$\hat{v}^{(m)}(q) = \hat{y}^{(m)}(q)\Lambda^{-1}(q), \quad 1 \le q \le N_g \tag{3.29}$$

Cancel ISI

$$\hat{y}^{(m+1)}(q) = \hat{y}^{(1)}(q) - \Re(H^{*}_{eq-qs}(q) H_{eq-nqs}(q))\hat{v}^{(m)}(q), \quad 1 \le q \le N_g \tag{3.30}$$

Form $\hat{\hat{x}}^{(m)}(q)$ from,

$$\hat{\hat{x}}^{(m)}(q) = A\hat{v}^{(m)}(q), \quad 1 \le q \le N_g \tag{3.31}$$

Cancel ICI

$$\begin{aligned} \hat{y}^{(m+1)}(q) = \hat{y}^{(m+1)}(q) - \Re(H^{*}_{eq-qs}(q)\bar{R}(q)\hat{\hat{x}}^{(m)}(q)) \\ - \left( H^{*}_{eq-qs}(q) \sum_{b=1, b\neq q}^{N_g} \bar{Q}(q,b)\hat{\hat{x}}^{(m)}(b) \right), \\ 1 \le q \le N_g \end{aligned} \tag{3.32}$$

m=m+1

*goto Loop* $z_1$.

It is noted that the above cancellation algorithm has polynomial complexity. Also, since $\wedge(q) = \Re\left[H^{*}_{eq-qs} H_{eq-qs}\right]$ is a diagonal matrix, its inversion is simple. In practice, accurate estimation of the channel coefficients is essential, which can be achieved, for example, using the algorithm proposed in [7].

## IV. SIMULATION RESULTS

We evaluated the SIR and BER performance of the proposed PIC detector for SFBC-OFDM in frequency and time selective Rayleigh fading channel through simulations. First, we illustrate the improvement in the output SIR achieved using the proposed IC detector (for the case of L=4 paths). We compare the output SIR performance with and without interference cancellation. With interference cancellation again has two cases i) only ISI (bypassing the ICI cancellation part in the algorithm), and ii) both ISI and ICI. From Figure 9 we observe that in time flat channels cancellation of ISI only improves performance significantly, and cancelling ICI also in this case does not bring additional benefits. But, for non-zero channel delay spread or user velocities, cancellation of ISI only is not adequate, and there is further improvement, when ICI is cancelled as well. For example at a speed of 30 Km/h, cancelling only the ISI improves the SIR by just about 4dB compared to no cancellation, whereas cancelling both ISI and ICI improves the SIR by about 20 dB for m=2.

In all our performance results in this section, we have used a 5.4 GHz SFBC-OFDM system having $N_c$ = 64 subcarriers with a subcarrier spacing of 312.5 KHz, and Nr=1. A frequency selective tapped-delay line channel model with L = 2 and L = 4 having equal powers is used. The codes considered are Alamouti’s [2]. We plot the BER performance of the proposed PIC detector in time-flat, frequency selective Rayleigh fading with L = 2 and L = 4 equal power paths for

SFBC-OFDM using Alamouti's code and BPSK/QPSK/16- QAM is shown in Figure (7- 9). So there is no time selectivity induced ICI here. However, there is frequency selectivity induced ISI. In Figure 9, Stage-1 performance corresponds to the case of no cancellation, where as Stage-3 performance is after two stages of proposed cancellation. For the sake of comparison, we also plotted the performance for time -flat and frequency-flat fading (i.e., no ISI and no ICI), which provides the best possible performance.

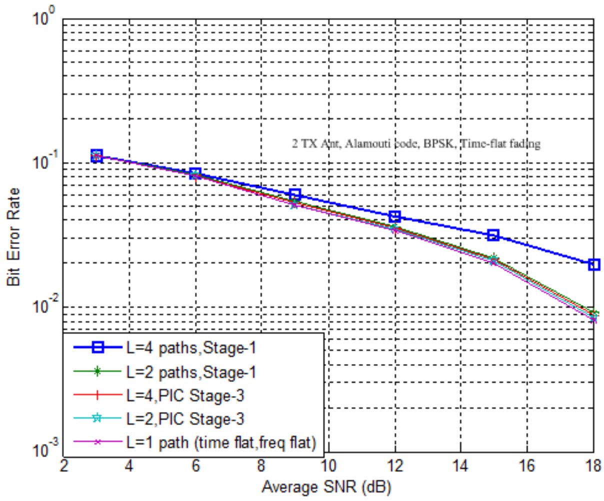


**Figure 7: BER Performance of the Proposed PIC Detector in Time Flat, Frequency Selective Rayleigh Fading (BPSK)**

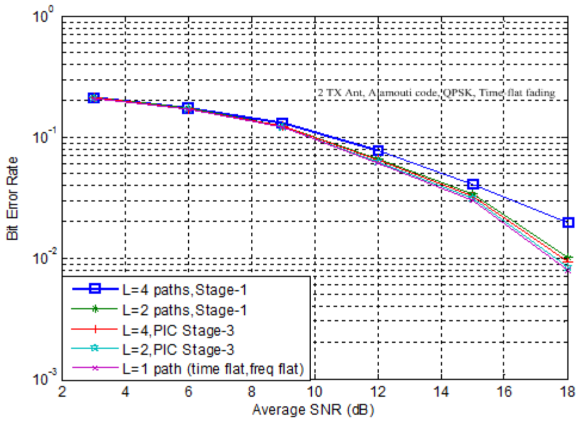


**Figure 8: BER Performance of the Proposed PIC Detector in Time Flat, Frequency Selective Rayleigh Fading *(QPSK)***

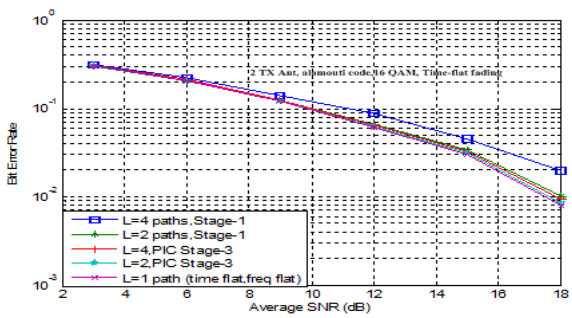


**Figure 9: BER Performance of the Proposed PIC Detector in Time Flat, Frequency Selective Rayleigh Fading (16-QAM)**

From Figure 9, it can be observed that for L = 2 (i.e., channel delay spread is small), The ISI induced is small, and hence there is no major performance improvement due to the cancellation. However, for L = 4 (i. e, delay spread of the channel is large), the ISI induced is high, and, in this case, the proposed cancellation results in significant performance gain (e.g., about 2 dB gain at $2x10^{-2}$ BER).

## V. CONCLUSIONS

In this paper, we investigate The BER performance of the SFBC-OFDM system using BPSK modulation is better than the BER performance of using QPSK modulation by approximately 3 to 4 dB. In this paper, we presented an interference cancelling algorithm for cancelling frequency selectivity induced ISI and time-selectivity induced ICI in SFBC-OFDM systems. In the first step of the algorithm, an estimate of ISI is obtained and cancelled, and in the second step an estimate of the ICI is obtained and cancelled. This two-step procedure is repeated in multiple stages to reduce the ISI-ICI induced error-floors. The presented simulation results show that the proposed detector effectively mitigates the effects of ISI and ICI in SFBC-OFDM systems. The proposed PIC detector algorithm can be easily extended to space-time-frequency (STFC) coded OFDM as well.

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